MEASUREMENTS ON MODERN VHF/UHF FRONT-ENDS

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IN MY earlier article "Modern vhf/uhf front-end design", [1] I explained how to design a receiver front-end with a specific noise figure and optimized gain distribution. So you design a new front-end, build it, and get it going. And then what? It may seem *to be working well—but is it, really? On-the-air impressions can be very misleading, unless you already have some idea of the kind of performance to expect. If you want optimum performance from a front-end, you need to check it out by* measurements.

This article describes some of the tests you can make on vhf/uhf front-ends or similar rf systems. Ideas and circuits are included for a variety of homebuilt testgear, with an emphasis on methods and equipment which give accurate results without needing external calibration.

Part 1 is about basic principles and the measurement of gains, losses and vswr. Part 2 will be about the measurement of sensitivity and strong-signal handling.

Part 1

ANY KIND of test or measurement involves three aspects: the device under test, the testgear, and yourself. If you have designed the circuit thoroughly in the first place [1], you should have a pretty clear idea how each stage should perform. If the results fail to agree with your expectations, there *must* be a logical reason, and it must lie in one of those three areas: the circuit, the testgear, or your own understanding of what's going on. You and the testgear are supposed to be testing the circuit, though it's quite possible that you and the circuit will finish up testing the testgear. The trick is to avoid the circuit and the testgear ganging up against *you*!

Modern commercial rf testgear is amazing and wonderful, but hideously expensive. Given the choice, most radio amateurs would prefer a new car or a new house instead! But don't despair; many of the features that make commercial rf testgear so expensive are unnecessary for amateur radio. For a start, we rarely need broadband performance across a wide frequency range. The broadband performance of commercial testgear always involves some degree of compromise, and home-made testgear that has been designed and optimized for a specific amateur band will often work just as well—or even better.

We don't need high accuracy in amateur radio, since we have no formal performance specifications to meet, and we can often manage with relative rather than absolute measurements. For example, it is easy to make a relative measurement of receiver sensitivity using an uncalibrated noise generator, but in order to make an absolute sensitivity measurement you would need to measure the receiver noise temperature or noise figure—a far more difficult project, as we shall see in Part 2.

Simple measurements, based on the calibrations of voltage and current meters, or on the marked values of components, can take you a surprisingly long way into the measurement of gains, losses, power levels and even strong-signal handling. You can also calibrate your home-made testgear against commercial instruments. The test and measurement stands at conventions are there to help you. One way or another, you can usually achieve the accuracy you need. Even so, the more your measurement methods wander away from reliable basic standards, the more errors can accumulate. The best test methods are the ones that calibrate themselves.

Gain and loss measurements

Power gains and losses are the basic currency of front-end system design. It's important to be able to measure them reasonably accurately.

The gain or loss of a device is the power level at its output, relative to the signal level fed into the input [2]. To measure gain or loss, you need a signal source, and an instrument for measuring *relative* power, calibrated in decibels. Although you don't need to know the absolute power level (in watts), you do need to make your measurements at the right *sort* of power level. To make valid measurements on receiver front-ends, you usually need to work at power levels well below the gain compression point [1], ie at levels of tens to hundreds of microwatts. At this low level, accurate relative power measurements require a special instrument which I'll describe later.

Say you want to measure the gain of an rf amplifier. First, set up the measuring system (Fig 1) with a coaxial adaptor instead of the amplifier. The two 20dB attenuator pads are used to establish 50Ω source and load impedances for the amplifier under test; see below. Adjust the meter so that it reads 0dB, then remove the adapter and insert the amplifier. The meter reading increases, and the amount of increase (in decibels) is the gain of the amplifier. If you had inserted a long length of 50Ω coaxial cable instead, the meter reading would have fallen below the 0dB setting, and you'd have measured the cable loss. As well as simple losses and gains, you can use the same technique for practically any other measurement that involves relative rf power levels, from hf to microwaves. For instance, you can measure filter responses, crosstalk in diplexers and coaxial relays, antenna gains and radiation patterns, and the vswr of anything that's got a vswr.

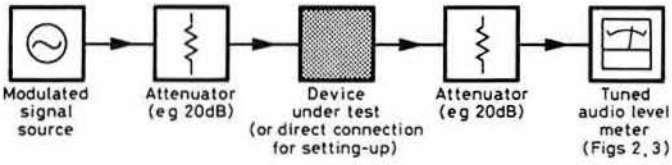
Modulated signal source
Attenuator (eg 20dB)
Device under test (or direct connection for setting-up)
Attenuator (eg 20dB)
Tuned audio level meter (Figs 2, 3)

Fig 1. Typical test set-up for gain or loss measurement. Note the use of attenuators at the input and output of the device under test

Stage-by-stage measurement of gains and losses is an extremely powerful technique for checking out a newly-built rf system such as a receiver front-end. If you've designed the system properly [1] you'll already know what to expect. This is one of the big rewards for spending time on the design before picking up the soldering iron; if there are problems, you're almost bound to spot them! It's always useful to design and build systems with 50Ω interconnections, so that you can test each stage separately. You don't need to go to the extreme of building each stage in a separate box with plug and socket connections. Even a single-board layout can be designed to include interstage connections of 50Ω stripline with in-line coupling capacitors. By removing a capacitor you can attach a coaxial test lead to the output of one stage or the input of the next. Thus you can step through the entire system, stage by stage, checking that all the gains and losses agree with the design values.

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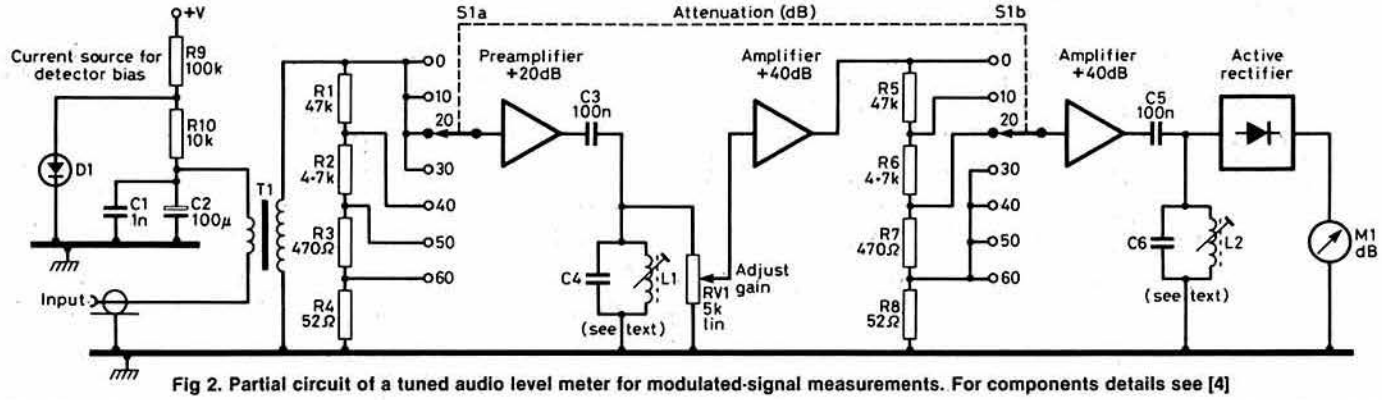

Fig 2. Partial circuit of a tuned audio level meter for modulated-signal measurements. For components details see [4]

Attenuators

One way to calibrate your relative power measurements is by comparison with fixed resistive attenuators. You'll need to build or acquire a range of good-quality attenuators designed for use in a 50Ω system. These should include a switched set giving any combination of 1, 2, 3, 4, 10 and 10dB (so you can select 1 to 30dB in 1dB steps), together with fixed attenuators of handy values like 3dB, 6dB, 10dB and 20dB. You can often pick these up as surplus, though home-made attenuators can be reasonably accurate at vhf, and with care can even work well into the uhf region [3]. It's also useful to own or have access to a few attenuators of known high quality, to check the others against.

Almost all your measurements are going to be made in a 50Ω system, feeding the device under test from a 50Ω source impedance and terminating it in a 50Ω load. You can ensure this by placing the device under test betweeen two 50Ω attenuators of 20dB or so. Any power reflected from a mismatch at the far side of an attenuator will itself be attenuated on the way back, so the vswr seen by the device under test can't stray far away from 1:1. For example, the vswr looking into a 20dB attenuator theoretically cannot exceed 1·02, no matter what is connected at the other end. (In practice the vswr would probably be rather higher, owing to errors in the attenuator itself.) If you make all your gain and loss measurements using impedance-stabilizing attenuators at the input and output of the device under test, you'll know where you are. If you don't, you'll get confusing and misleading results. For example, even your so-called standard attenuators won't perform as designed unless they themselves are in a 50Ω system.

Tuned af level meters

As I mentioned earlier, gain and loss measurements on receiver front-ends need to be made at power levels of tens to hundreds of microwatts. Conventional power meters using diode detectors and dc voltmeters either aren't sensitive enough, or tend to suffer from noise and drift. A better solution is to use a test signal which is amplitude modulated by a steady audio tone, and to measure the relative level of the *tone* instead of the carrier. The signal is detected by a diode in the usual way, but instead of trying to dredge the dc component out of the hum and noise, you amplify the audio tone signal. AC-coupled amplifiers neatly sidestep the dc drift problem, and hum and noise are reduced by a sharp audio filter tuned to the tone frequency. Finally you measure the level of the tone on a meter calibrated directly in decibels, those being the units of relative power level.

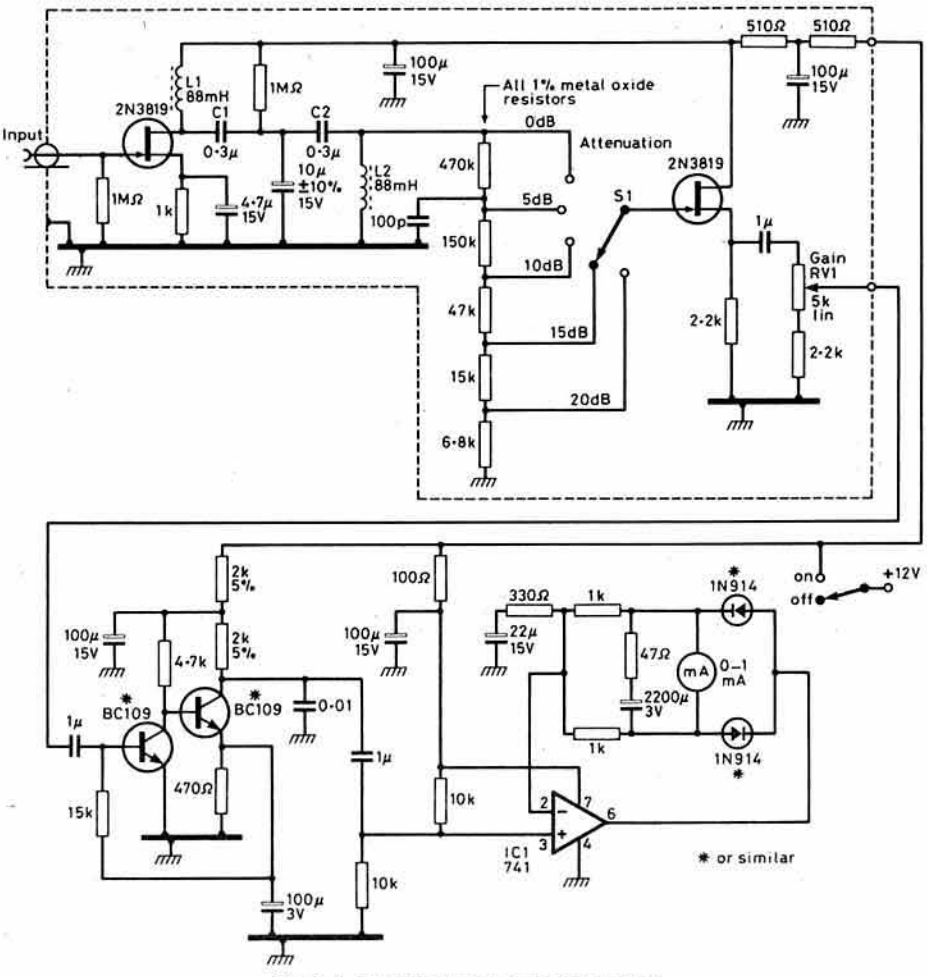

Fig 3. A simplified tuned audio level meter

Testgear for the modulated-signal technique is very easy to build. The only special instrument is the meter which measures the level of the audio modulating tone. Fig 2 shows a partial circuit diagram [4]. It's simply a tuned af amplifier with adjustable gain, followed by an active rectifier which drives a meter. Gain is adjustable by both a switched attenuator and a continuously-variable control. The switched attenuation is in 10dB steps; the total of 60dB is split into two separate 30dB attenuators to maintain a good signal:noise ratio without overdriving any stage.

My own instrument is built to the same principles as Fig 2, but it's now 10 years old and is beginning to look rather antique with all its discrete transistors. You could easily update the instrument to use modern low-noise op-amps, so long as you keep to the basic block diagram.

Construction should follow standard hi-fi preamplifier practice. The first 20dB gain block is a low-noise audio preamplifier and requires care with shielding and grounding around the input. The two 40dB gain blocks are much less critical. Any frequency around 1kHz will do for the two tuned circuits, as long as they're both the same. The inductors L1 and L2 in my instrument are two pot cores from the junk box; the components suggested in [4] are the nearest commercially available. Capacitors C3 and C5 are chosen to match the low output impedances of the gain blocks to the high-impedance tuned circuits; the values shown should be satisfactory, though some adjustments may be necessary.

Fig 3 shows an alternative instrument [5] which uses the well-known 88mH inductors for selectivity

To get the best from either instrument, you need a large, accurate, non-digital meter with some existing form of linear calibration (you'll see why in a moment). The instrument should preferably be battery powered, to avoid direct pickup of the modulating signal via ground loops instead of via the device under test.

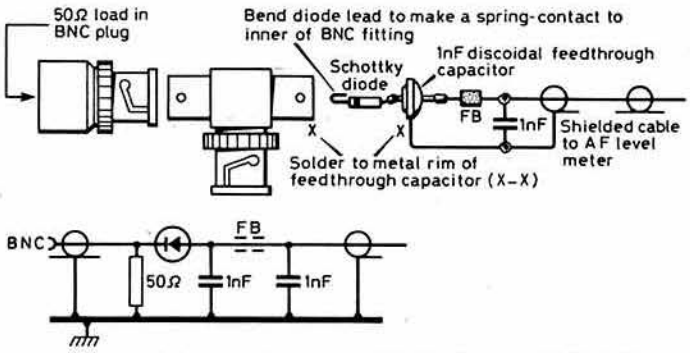

Fig 4. Exploded view of a detector and 50Ω load using a BNC T-fitting

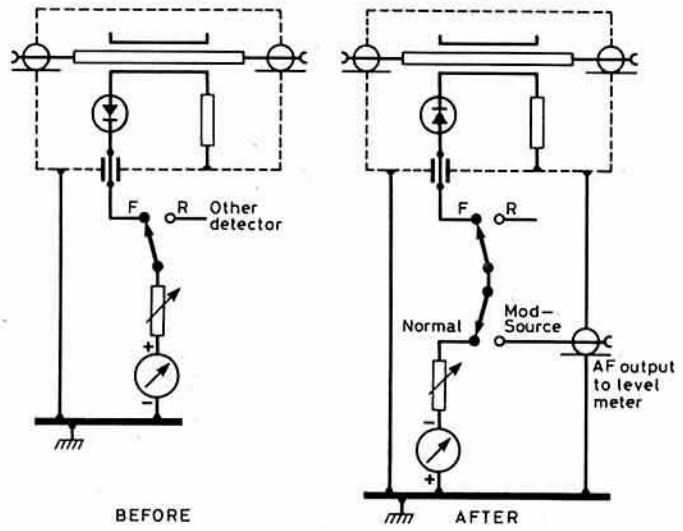

Fig 5. Adapting an ordinary "vswr bridge" for modulated-signal measurements. The two detector diodes in the bridge are reversed, as are the meter connections

Detectors and vswr bridges

A variety of diode detectors can be used for the modulated-signal technique. The most convenient are ordinary Schottky diodes fed with a few microamps of dc forward bias provided by the level meter (Fig 2). At or below the milliwatt rf level, these diodes are in their square-law region, which means that the rectified audio level is accurately proportional to the rf *power* (not the voltage). If your meter scale already has an accurate linear calibration, you can add a decibel scale using nothing more than a calculator [6]. Detectors usually need to provide a 50Ω rf termination, and I use a Schottky diode in a BNC tee adaptor with a 50Ω BNC load attached (Fig 4). The rf impedance is not exactly 50Ω, so I also use an attenuator ahead of the detector to "flatten" its vswr.

Since the modulated-signal technique basically measures power ratios, you can also measure vswr. Although vswr is defined as a ratio of impedances, it is more often measured as a ratio of the "forward" and "reverse" rf power levels or voltages, as detected by directional sensors. An ordinary vswr meter contains these directional sensors but typically requires at least a watt of rf to give a good indication on the dc meter display. Simply reversing the diode detectors allows you to use the modulated-signal technique to measure the vswr of delicate circuits at very low power levels (Fig 5). For example, you can measure the input or output vswrs of preamplifiers, mixers and other allegedly 50Ω devices (prepare for nasty surprises!).

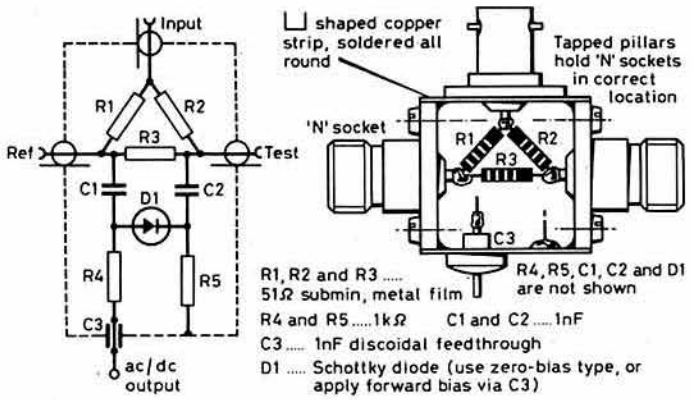

Fig 6. Circuit and sketch of a return loss bridge

Other kinds of vswr sensors can also be used. Fig 6 shows a home-made vhf/uhf return loss bridge. Return loss is an alternative way of expressing vswr, and can be read directly in decibels from the meter scale [7]. W7ZOI has recently described many uses for an hf/vhf test set based on an hf-type return loss bridge and unmodulated rf sources [8]. By using the modulated-signal technique and the bridge of Fig 6, the same measurements can be extended to uhf. For microwaves, you can use a slotted coaxial line or waveguide to observe and measure standing waves more directly [9].

Modulated signal sources

You can use a wide variety of modulated signal sources with the same detectors and level meter. For example, on 432MHz I use a crystal-controlled source modified from a converter local oscillator strip, which provides a few milliwatts of rf. A "Microwave Committee" board [10] would serve equally well if throttled back to the same power level. Since any signal source will always need a 50Ω output attenuator to establish the correct source impedance, a suitable pad can be permanently built in. Modulation can be very simple; almost any waveform will do, so long as the frequency is adjustable to match the tuned circuits in the level meter. My 432MHz source is modulated by an af oscillator using a 555 ic, which chops the rf output by supplying square-wave forward bias to the base of the final transistor—a bit brutal, but very effective! Whatever kind of modulated source you use, it should be very well screened, with supply leads decoupled to avoid stray pickup into the detector. The rf output should also be spectrally pure, because the device under test may have significant—and misleading—responses to spurious frequencies from the signal source.

Modulated signal sources can be made for all the vhf/uhf amateur bands of interest, following the above guidelines. The modulated-signal technique can also be used outside the amateur bands, and is equally useful at hf. For example, you could use a suitable general-coverage a.m signal generator with its internal modulation adjusted to the peak audio response of the meter. This would allow you to measure gains and losses in the i.f stages of a front-end, and to manually sweep the frequency responses of filters. If you are sweeping over a wide frequency range, or are measuring the gain or loss of a frequency-translating device (eg a mixer, transverter or complete front-end) you are also relying on your detector having a flat frequency response. It would be wise to check the detector first, using a good signal generator.

Accuracy

The modulated-signal technique can measure gains and losses with excellent accuracy. From the 0dB reference at full scale on the meter, the first 1dB of loss is spread over 21 per cent of the meter scale [6] so you can easily resolve changes of less than 0·1dB. Changes of 10 or 20dB are taken care of by the range switch, while bigger changes require external rf attenuators. Since the calibration of the instrument relies only on resistor values and the linearity of the 0—10 meter scale, you can use it to cross-check the calibrations of your attenuators. Take care to make all measurements at power levels within the square-law region of the detector diode. You can check this by increasing the applied power by a known amount, and observing that the meter reading increases by exactly the same amount. All the calibrations—the meter scale, the 10dB range switches, and all your fixed and switchable rf attenuators—are capable of being consistent within a small fraction of a decibel [11], at frequencies up to at least 432MHz.

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opinion anyone using headphones—no matter what the state of their hearing—should use some form of automatic peak audio limiting, such as that provided by a pair of back-to-back diodes. Reg Taylor, G3AVQ, draws attention to yet another reason why you should avoid overloud noises near your ears. He writes:

"May I suggest a further warning. Even if loud signals do not impair or destroy hearing they can cause 'tinnitus', otherwise known as 'ringing in the ears' or 'noises in the head'. Being a sufferer myself since May 1979, I did for a while act as a research patient for the Royal National Institute for the Deaf. I learned then that not only is excessive noise one cause of tinnitus but that it may not manifest itself until years after exposure to the noise (mine could have been caused by second world war gunfire and there is, I believe, an association of wartime air gunners who suffer from it). I have read that it affects as many as 17 per cent of the population, a figure that may well go up dramatically when the effects of the Walkman-type portable tape cassette machines begin to show! In my case, tinnitus takes the form of a constant, 24h/day, 7kHz purish tone at a level of S4 to S5 (on a 1 to 9 scale). For others the frequency may be in another part of the audio spectrum and may be intermittent rather than continuous. Whatever form it takes I can assure *TT* readers that it is something to be avoided if possible."

The 10·1MHz band

A considerable number of readers have commented, some constructively others abusively, on the suggestion (*TT* August) that it is time to consider whether a small segment of the narrow 10·1MHz band might logically be used for ssb, if only to encourage more amateurs to use this valuable band than do at present. I have to admit that virtually all those who took the trouble to write to me were highly critical of this proposal and of the technical justifications put forward by G6XN. While, personally, I have no wish to use ssb on this band, I make no apologies for providing a forum for G6XN's views, remembering that among many other feats, he was the first person in this country to draw attention to chordal hop.

Several correspondents have pointed out that such discussion is a little academic, since the main reason for the under-utilization of the band has always been the presence of so many high-power "primary users", ie the fixed (point-to-point) commercial service. The amateur allocation is strictly on a secondary, non-interference basis. It remains a paradox that for much of the time the spectrum just below 10·1MHz or just above 10·15MHz seems far less crowded with commercial signals than 10·1 to 10·15MHz! The 1986 edition of Joerg Klingenfuss's *Guide to Utility Stations* lists no less than 19 fixed-service stations between 10,100 and 10,150kHz. These include cw, rtty, fax and ssb stations, many of them concerned with the meteo (weather) service, and many putting strong signals into the UK for much of the time.

One of the problems is that commercial stations, in order to keep their channel open, often spend hours "idling". It has been shown by various professional investigations that in Europe after dark it is unusual, at least in sunspot minimum periods, to find, below the muf, *any* interference-free gaps exceeding about 200Hz.

David Sergeant, G3YMC, one of those firmly against any use of ssb on this band, points out that since it is a secondary allocation to amateurs there is little likelihood that the long-established high-power rtty, fax etc commercial stations will move out in the foreseeable future. This means that in reality only about half the band, or about 25kHz, is actually usable by amateurs. He feels that if ssb stations were encouraged to use the top end of the band, they would soon be forced to spread downwards until they occupied virtually the entire available spectrum. He accepts, nevertheless, that the band is seriously under-utilized at present. He believes that more cw activity would be generated by accepting 10·1MHz contacts for "award" purposes—though, like others, he feels strongly that the band should continue to be kept clear of "contest" operation.

Several readers doubt whether we are ever likely to see the use of notch and narrowband filters effective enough to permit spectrum sharing between cw and ssb. Some dismiss the very idea as nonsensical; my own opinion is that cw operators might come off better than ssb operators in such a situation!

Phil Stevens, G3SES, a regular user of 10·1MHz, notes that one reason for the lack of cw activity is the continued absence of Russian stations that provide so significant an element of the cw activity on 7, 14 and 21MHz. He also believes that the widespread use on other bands of trapped dipoles and trapped verticals not designed for 10·1MHz discourages use of the band. His more provocative suggestion (no letters to me about this *please*) is that the Class B licence should be scrapped and replaced with a cw-only 25W "novice" licence for 1·8, 10, 18, 24 and 28MHz. In his view this would ensure a better standard of cw operation and more use generally of hf. □

MEASUREMENTS ON MODERN VHF/UHF FRONT-ENDS

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References and notes

[1] "Modern vhf/uhf front-end design", Ian White, G3SEK. *Rad Com* April-July 1985, pp264, 367, 445, 537.

[2] Strictly speaking, we are measuring *insertion* gains and losses. Regardless of the actual input and output impedances of the device under test, the input power is defined as the power available into a 50Ω load, and the output power is that delivered into a 50Ω load.

[3] Design data on home-made attenuators for vhf/uhf are given in *Microwave Newsletter Technical Collection* (RSGB); the same data are in Appendix I to "The effect of preamplifiers on receiver performance", J N Gannaway, G3YGF. *Rad Com* November 1981, p1026.

[4] The original source of the circuit in Fig 2 was an article by W6VSV in a pre-1970 edition of *Ham Radio*. In the USA such an instrument is known as an "swr meter", a hangover from its original use as an indicator for a slotted-line probe. Hewlett-Packard still market the HP415 under this name, which conceals the instrument's far wider uses.

Components details:
D1 Silicon, eg 1N914;
L1, 2 80–100mH, eg Toko 719VXA8032 80mH (Cirkit Ltd);
R1–8 1% metal oxide;
R4, 8 47Ω + 4·7Ω;
T1 1:5, eg "transistor interstage" type;
C4, 6 220nF approx to resonate at 1kHz with L1,2.

[5] "Antenna gain measurements", Fred Brown, W6HPH. *QST* December 1982, p27. The article also describes simple tunable modulated signal sources for 432MHz and 1·3GHz, and tuned diode detectors which may give useful rejection of unwanted external signals when testing antennas on an open-air range. The 88mH inductors used in Fig 3 are available to UK amateurs through the G-QRP club.

[6] The scale calibration of a square-law diode detector is given by: (fraction of full-scale deflection) = 0·1 antilog (decibels below full-scale). If the full-scale reading of 10·0 is marked as 0dB, the rest of the scale can be marked as follows:

Loss (dB)	Scale (0–10)	Loss (dB)	Scale (0–10)
0·0	10·00	6·0	2·51
1·0	7·94	7·0	2·00
2·0	6·31	8·0	1·59
3·0	5·01	9·0	1·26
4·0	3·98	10·0	1·00
5·0	3·16	20·0	0·10

Note how this non-linear scale is expanded where you need the best resolution, in the region of low losses. The first decibel of loss is spread over more than 20 per cent of the entire meter scale.

[7] "More about return loss bridges", Ian White, G3SEK. *The Lunar Letter*, May 1983. VSWR and return loss L (in decibels) are related by the formula

$$\text{vswr} = \frac{1 + \text{antilog}\ \frac{10}{10}\ \frac{(L/20)}{(L/20)}}{1 - \text{antilog}\ \frac{10}{10}\ \frac{(L/20)}{(L/20)}}$$

A high return loss indicates a low level of reflected power, and hence a good vswr. A perfect open or short circuit has zero return loss. To use a return loss bridge, first connect a *good* 50Ω load on the REF port. Set the af level meter to zero with an N-type short-circuit connected to the TEST port, then connect the device under test and simply read its return loss from the meter.

[8] "Beyond the dipper", Wes Hayward, W7ZOI. *QST* May 1986, p14.

[9] "How to use an swr indicator", Bob Stein, W6NBI. *Ham Radio* January 1977, p66.

[10] *Microwave Newsletter Technical Collection* (RSGB)

[11] The accuracy of this method is quite amazing, considering that everything is home-built. In my own set-up, external 10dB attenuators correspond to 10dB steps on the range-switch within a meter needle's thickness. The meter display is the least accurate part, but only because of non-linearity in the meter movement itself! That problem could have been avoided by ignoring the existing meter calibration, and marking the decibel points with the aid of an external digital instrument.

TO BE CONCLUDED

MEASUREMENTS ON MODERN VHF/UHF FRONT-ENDS

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PART 2

Noise measurements

Measurements related to noise are another stock-in-trade of front-end performance testing. In the test methods I shall describe, the test signal itself is a known amount of broadband rf noise, and we are interested in measuring the relative difference in noise levels when the test signal is applied. Noise levels are most conveniently measured at the audio output of the receiver. When you do this, you are relying on the audio output being strictly proportional to the noise or signal level, so the receiver's agc must be switched off for all the tests to be described. You then need to be careful to avoid limiting or distortion in any rf or af stage [1]. *Always repeat your measurements,* especially where random noise is involved. Repeat them under the same conditions, to get some idea of the random errors. Then alter the rf and af gain settings and repeat them all again. If the results change, find out why.

A suitable instrument for relative af noise power measurements is a rectifying meter as in Fig 7, which can be pre-calibrated in decibels by calculation [2]. This is essentially a peak-detecting instrument, so you need to be particularly careful that noise peaks aren't being significantly clipped by overloading of any stage in the receiver. More complex rms-detecting audio noise level meters are less prone to this error, but with care the simple meter in Fig 7 is quite adequate. With such a meter you can accurately measure (signal + noise)/noise ratios, as required in any of the following techniques for sensitivity measurement.

Sensitivity measurements

VHF/UHF receivers are extremely sensitive. As I explained in the previous article on front-end design [3], a good 144MHz front-end would have a noise figure of about 2dB, and would be able to detect signals as weak as 20 *nano*volts. At those levels, accurate sensitivity measurements using an attenuated signal generator are extremely difficult. The signal-generator technique is still satisfactory for hf receivers and relics from the older generation of deaf vhf/uhf receivers; but accurate measurements on low-noise systems require an extremely well screened signal generator and meticulous attention to both practical and theoretical details. For modern vhf/uhf front-ends, better techniques for sensitivity measurement are those that stay as close as possible to the fundamental concept of noise temperature [3]. Best of all, measure the noise temperature itself.

Noise temperature measurement techniques are based on supplying a known amount of excess noise to the rf input of the receiver, and measuring the change in noise output. If the receiver noise temperature to be measured is T_{RX}, and the noise source has a noise temperatures T_{ON} when switched on and T_{OFF} when switched off, the system noise temperatures are:

$$T_{SYS} = T_{RX} + T_{ON} \quad \text{(noise source on), and}$$
$$T_{SYS} = T_{RX} + T_{OFF} \quad \text{(noise source off).}$$

The noise power from the receiver is proportional to the system noise temperature, so the on/off power ratio Y is given by:

$$Y = (T_{RX} + T_{ON})/(T_{RX} + T_{OFF})$$

If we know T_{ON} and T_{OFF}, and can measure Y, we can calculate T_{RX}.

$$T_{RX} = (T_{ON} - YT_{OFF})/(Y - 1)$$

The hot/cold method

The main problem is to generate rf noise at two different known noise temperatures T_{ON} and T_{OFF}. The simplest way is to use the thermal noise from a 50Ω resistor at two known physical temperatures (Fig 8(a)). This is called the hot/cold method, and is the closest you can get to a fundamental measurement of noise temperature. For the best accuracy you need a large difference between the two temperatures. Resistors don't much like being roasted, so the hot end of the range can't go too high. The cold end can be extended down to 77K if you can get liquid nitrogen from your friendly local laboratory (in rural areas, ask the AI man), or to about 196K in crushed solid CO_2. Unless you know how to handle these potentially hazardous materials properly, you'd be safer to use a shorter temperature baseline, eg between melting ice and hot water. With care and a good laboratory thermometer, you can still obtain quite accurate results.

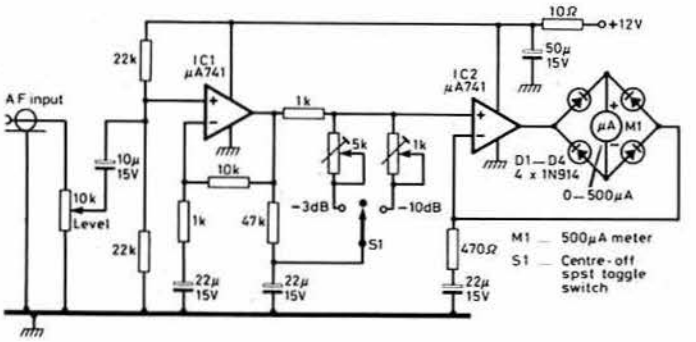

Fig 7. A peak-responding audio voltmeter for use in measuring (signal + noise)/noise ratio. For full details see [2]

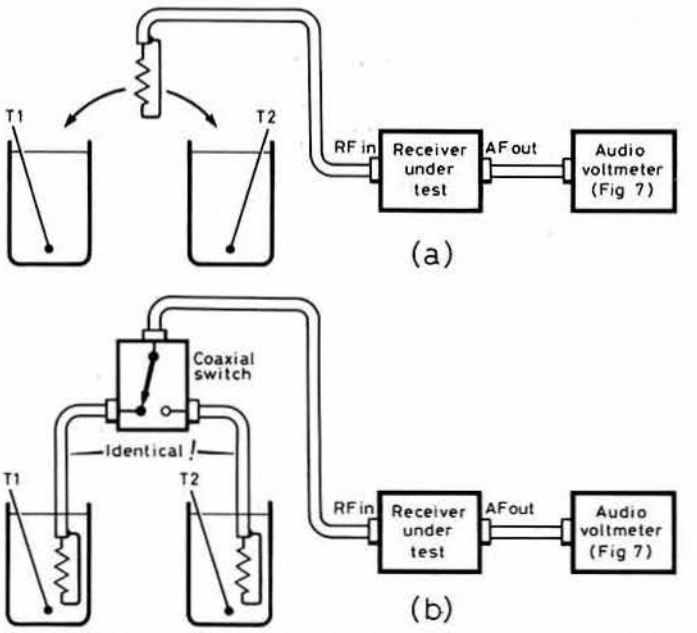

Fig 8. The hot/cold method for measuring noise temperature. The basic method (a) involves transferring one resistor between two temperature-baths. The two-resistor method (b) is more convenient but requires even more care in setting up

Even the fundamental and basically simple hot/cold method has pitfalls for the unwary. To present an accurate 50Ω termination at both temperatures, the resistor must have good rf characteristics and a low temperature coefficient. Corrections need to be made for losses and thermal noise from connecting cables. There may be errors owing to drift in system gain while the resistor is being warmed up or cooled down, and measurements can become very tedious and slow. It's far quicker to compare noise powers from two 50Ω resistors at different temperatures (Fig 8(b)), but then the two resistors and all their associated connections need to be identical in all respects (except temperature) so the test system itself requires careful preliminary checking before the results can be trusted.

Non-thermal noise sources

Another way of generating noise is to use a non-thermal source, such as the noise generated in certain kinds of diode. The traditional method was to use a thermionic diode or a silicon microwave detector diode in forward conduction. Nowadays a better and cheaper method for amateurs is to use a diode in reverse breakdown, eg a zener or the base-emitter junction of a silicon transistor. With careful control of diode current, the long-term noise characteristics can be quite repeatable.

It is absolutely vital to make sure that the noise generator represents an

* 52 Abingdon Road, Drayton, Abingdon, Oxon OX14 4HP.

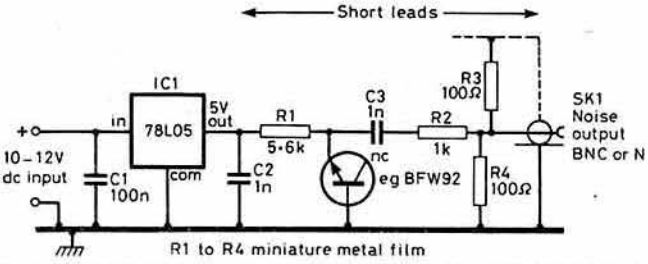

Fig 9. A diode noise generator with low vswr in "on" and "off" states. The circuit from C2 to the output socket SK1 is built with the shortest possible leads on a copper groundplane (pc board) soldered directly to the back of SK1. R1 may be altered to maximize noise output

accurate 50Ω source, and the impedance must not change when the generator is switched on and off. Many published noise generator circuits ignore this precaution. The noise temperatures of many modern preamplifiers (especially GaAsfet types) are extremely sensitive to source impedance, and measurements can be totally falsified by changes in the on/off impedance of the noise source. A suitable diode noise generator is shown in Fig 9. The diode is a transistor base-emitter junction operated in reverse breakdown, the resistor in series with the regulated 5V supply having been selected to maximize the noise output. The on/off impedance change is stabilized by over 30dB of rf attentuation; the rf attenuator arrangement in Fig 9 was optimized at 432MHz with the aid of a home-made return loss bridge (part 1, Fig 6) and its performance was subsequently verified using commercial equipment. Regardless of whether the dc supply is on or off, the noise generator presents a good 50Ω termination up to 432MHz, and is still passable at 1·3GHz. Although its noise temperature is not calibrated, the generator can be used in comparative measurements and for checking system performance. It can also be used with the G4COM noise figure comparator [4] which is highly recommended for aligning front-ends to give the best possible nf.

The problem with all semiconductor noise generators is that they can't be used for absolute measurements until their noise temperatures have been calibrated against some other standard, preferably a hot resistor. Even so, their convenience and repeatability has made semiconductor noise sources the norm in commercial noise generators for vhf, uhf and at least the lower microwave region.

One type of diode noise generator that is often claimed to be self-calibrating is the saturated thermionic diode, whose absolute noise output is theoretically predictable. But this is another example of a traditional method which doesn't work with today's low-noise front-ends. There's a hidden approximation in the normal procedure of measuring the diode current corresponding to a 3dB increase in $(s+n)/n$ ratio, which leads to serious errors with low-noise receivers [5]. Also, although the textbooks take account of the thermal noise from the 50Ω terminating resistor in the diode circuit, they usually assume that the resistor is always at 290K. In fact the resistor sits right next to a thermionic diode in a hot little box, and with a low-noise front-end you could *hear* that resistor warm up as the measurement session proceeds! Worst of all, the output impedance of most thermionic noise generators varies dramatically between the "on" and "off" states, and I've already mentioned that many modern low-noise amplifiers don't like that. For accurate measurements on modern vhf/uhf front-ends the thermionic diode noise source is no more self-calibrating than a semiconductor source, and a lot less convenient.

To sum up, relative measurements of noise temperature with uncalibrated noise sources are quite straightforward, given a certain amount of care and basic understanding. But there are no simple methods for absolute measurements. The hot/cold method is probably the most promising for use at home, though even this needs a great deal of care.

Strong-signal (dynamic range) measurements

As I explained in [3], strong off-frequency signals can have several possible effects on receivers. The following tests evaluate the three effects that are of most concern to amateur vhf/uhf operators because of their potential to interfere with weak signals. The three effects are:

***Third-order intermodulation**—mixing between at least two strong in-band signals, to give products which are also in-band and can interfere with the wanted signal.

***Reciprocal mixing**—raising of the receiver's apparent noise level, due to the noise sidebands of the receiver's own local oscillator.

***Gain compression**—when a strong off-frequency signal causes one of the stages in the receiver to limit, suppressing all other signals including the wanted one.

Different levels of strong signals are needed to provoke each of these effects. These levels can each be expressed in decibels above the receiver's noise floor, giving the dynamic range for the overload effect in question [3]. Remember, there is no single definition of dynamic range: a receiver has a separate dynamic range for each overload effect, and thus needs a separate test. Even slight overload could interfere severely with a barely-copyable weak signal. When testing receivers meant for amateur dx working, we usually measure so-called "spurious-free dynamic ranges", which relate to the situation where the particular overload effect just begins to be noticeable.

Strong-signal sources

Strong-signal tests require signal sources capable of delivering cw carriers at known power levels of several milliwatts, with low levels of spurious signals including noise sidebands. For gain compression measurements on modern front-ends the level of the single-tone test signal needs to be variable up to about 0dBm (1mW). A power level of about −30dBm is required to test for reciprocal mixing—and many present-day commercial vhf/uhf receivers are in deep trouble with test signals of much lower power!

Home-built signal sources for vhf and above need to be crystal-controlled, and designs for low-noise crystal oscillators are available. Fig 10 is a typical example of a low-noise oscillator for the frequencies in which we are interested [6]. The oscillator is based on a fet cascode amplifier with positive feedback through the pi-network C10-L2-C11. The overtone crystal bypasses the source of TR1 to ground at its series-resonant frequency, producing sufficient loop gain to permit crystal-controlled oscillation. The active devices in most oscillator circuits not only provide gain but also limit the amplitude of the oscillations, so the devices have to operate in a non-linear mode. This can lead to increased levels of noise sidebands. In the circuit of Fig 10 the amplifying and limiting functions are separated; D1 and D2 do the limiting, allowing TR1 and TR2 to run in linear Class A for lower noise. Other low-noise features include the use of power fets at high bias currents, and the generally high signal levels maintained around the oscillator loop. Note the rf chokes RFC1 and RFC3 in series with RV1 and RV2, the source bias resistors of TR1 and TR3. Without these chokes, RV1 and RV2 would become sources of thermal noise because their resistances are quite well matched to the circuit impedances prevailing at these points. (I'm not so sure about RFC2, though, because the low input impedance of TR3 swamps the gate impedance of TR1 and produces a good mismatch to thermal noise generated in 100kΩ resistor R3.) Constructional and setting-up details for this oscillator are given in [6].

A low-noise oscillator deserves low-noise amplifiers and frequency multipliers. The amplifier TR3 in Fig 10 is a power fet, running at high current in Class A. Higher-level amplifiers can also use power fets, or bipolar vhf/uhf power transistors [7], which should be

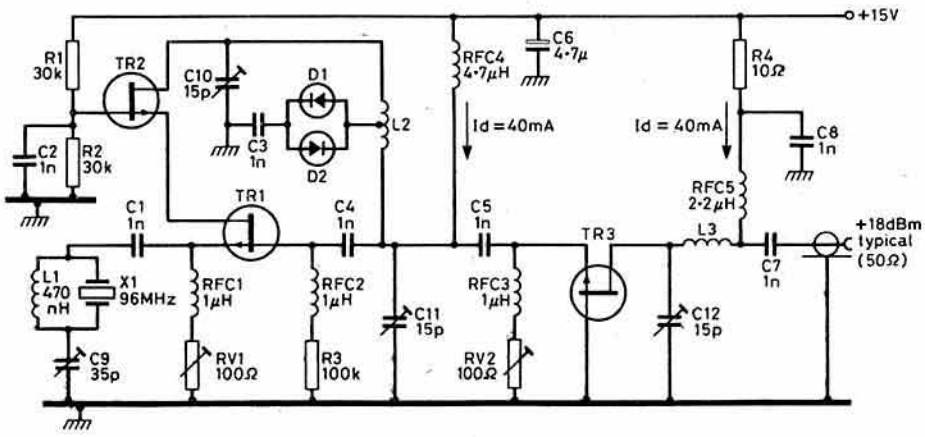

Fig 10. A low-noise crystal oscillator for 96MHz. The circuit is adaptable to 72MHz (× 2 = 144MHz) or 108MHz (× 4 = 432 MHz). For details see [6].
D1, 2: Schottky diodes
TR1, 2, 3: P8000 power fets (or U310 at Id = 20mA)
X1: 96MHz
L1: to resonate at 96MHz with X1
L2: 6 turns, 1mm diameter wire on 6mm diameter former, centre-tapped.
L3: as L2, no tap.

operated in Class A and considerably below their normal rf output ratings. Strong-signal sources for 50MHz and 70MHz can use overtone oscillators directly on those frequencies, while sources for the higher bands require frequency multipliers. Multipliers have to be non-linear, or else they wouldn't work, but the operating conditions of ordinary transistor multipliers are very vague indeed, making it hard to design for low-noise performance as well as reasonable efficiency. Avoid high-order multiplication in a single stage; frequency-double if you can, treble if you must, but nothing higher. Thus a 144MHz source requires an oscillator on 72MHz followed by a doubler (overall, this will probably be easier, cheaper and better than going for an oscillator on 144MHz direct), and 432MHz requires a 108MHz oscillator and two doublers. Following my own logic, a 1·296GHz source requires an 81MHz oscillator and four doublers, though I haven't tried it and frankly I'd be tempted to cut corners and frequency-treble at least once.

A frequency-doubling technique worth exploring is the balanced-diode circuit [8]. If the circuit in Fig 11 looks like a refugee from a power supply, that's because it actually *is* a full-wave rectifier. Remember how 50Hz mains input is converted into 100Hz ripple on the smoothed dc? In Fig 11 we apply an rf input at frequency f through a suitable centre-tapped transformer, let the dc output run away to ground, and use the rf ripple. This is mostly at the doubled frequency 2f, because the balanced arrangement tends to cancel out the neighbouring frequencies, f and 3f. A single tuned circuit is enough to give a very pure output at 2f. The diodes should be fast Schottky switching types for vhf/uhf, and for low noise and good doubling efficiency they need to be driven hard so they can't dither in partial conduction. Diode multipliers are passive, so they need a Class A amplifier at the output frequency to make up the loss in the diode circuit [8]. Comparing a diode multiplier chain with a conventional chain using transistor multipliers, you need about the same number of transistors, and you also need the diodes and their transformers. But diode multipliers don't need double-tuned circuits, and they give you a far better chance of obtaining low noise and good spectral purity without special setting-up. It's very satisfying to carefully design something, built it at home, tune it up "blind", and then take it into a lab and find it really *is* good!

Intermodulation measurements

Intermodulation requires at least two strong signals, so a test for intermodulation performance requires two separate signals fed into the receiver input. For accurate results, the only source of intermodulation between these two signals must be the actual receiver under test. This is tricky: somehow you have to bring the two signals together without letting them intermodulate. Intermodulation could occur in any non-linear components outside the receiver—most notably in the output amplifiers of the signal generators themselves, where signal levels are the highest. So the two signal generators need to be isolated from each other, by in-line attentuators plus a special hybrid combiner to add the two signals while keeping the sources isolated. Another attenuator after the combiner varies the level of the two-tone test signal. The basic test setup is shown in Fig 12, and more details are given in the references [7,9].

For measurements on modern vhf/uhf front-ends, the level of the two-tone test signal needs to be variable up to about −40dBm, with incidental intermodulation products suppressed by more than 60dB. (It is conventional to quote the levels of intermodulation products against the power level of *one* of the two equal test signals, rather than the sum.) Since the output of each signal generator has to pass through a high-value attentuator plus another variable attenuator after the combiner, each generator must provide an initial output of at least +10dBm. This means making *two* high-level low-noise signal sources following the guidelines I gave earlier. The direct rf signal paths are not the only source of intermodulation between two generators. One rf source can modulate the other via leakage along the *outside* of coaxial cables, or by rf getting on to the dc supply lines. Each signal source must be very carefully screened, and

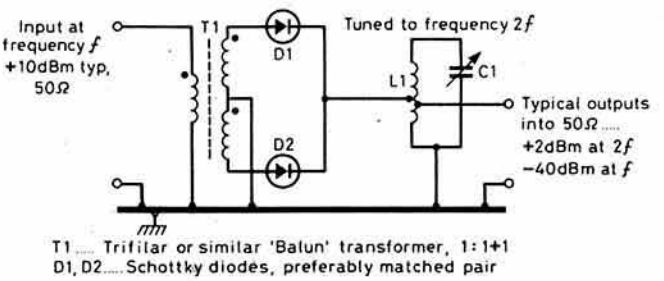
Input at frequency f +10dBm typ, 50Ω
T1
D1
D2
Tuned to frequency 2f
L1
C1
Typical outputs into 50Ω .... +2dBm at 2f −40dBm at f
T1 .... Trifilar or similar 'Balun' transformer, 1:1+1
D1, D2 .... Schottky diodes, preferably matched pair

Fig 11. Balanced-diode frequency multiplier for vhf/uhf [8]

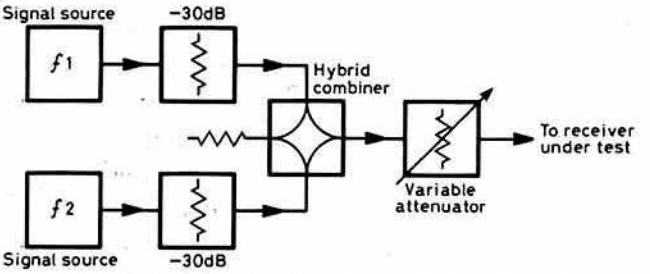

Fig 12. Intermodulation test set-up. Note the precautions to isolate the two signal sources from each other

all dc supplies must be fanatically filtered and shielded; typical precautions include boxes within boxes, each with soldered-up lids, double filtering of dc supplies through screened compartments, and permanently-soldered connections of solid-wall coaxial cable for the high-level signals [7, 9].

Having assembled the whole two-tone test set-up, take it to a convention where someone can put it on a spectrum analyser to check for unwanted intermodulation and to get an absolute calibration for the rf levels. You can calibrate everything by yourself, but why not make use of the facilities available? The people who run the test benches at conventions will be pleased to help—you'll be a welcome change from the procession of black boxes!

Conclusion

With the home-made test equipment I have described, you can make all the necessary tests and measurements on modern vhf/uhf receiver front-ends. Simple relative measurements may be all you need to optimize front-end performance, and to check that it stays that way. Beyond that, absolute measurements *can* be made with all the accuracy we need in amateur radio. It will take you longer than the person using professional test gear, but with care and understanding you *can* get good results.

What really matters is getting the best possible on-the-air performance from your receiver. *Any* measurement—no matter how simple—is better than guesswork and wishful thinking!

References and notes

[1] A common source of errors in measurements on receivers that have no front-panel agc on/off switch is to try to work with signal levels below the agc threshold, merely to save having to delve inside the set to disable the agc. Even if the S-meter needle does not move, the agc may still be operating on noise peaks, and your measurements will be useless and misleading.

[2] *Solid State Design for the Radio Amateur,* Wes Hayward, W7ZOI, and Doug DeMaw, W1FB (ARRL), p168. The response of this meter is proportional to the signal voltage (not power as in a square-law detector). The -3dB and -10dB presets can be calibrated by calculation from the meter scale, or by using fixed attenuators in a 50ohm audio system.

[3] "Modern vhf/uhf front-end design", Ian White, G3SEK. *Rad Com* April-July 1985, pp264, 367, 445, 537.

[4] "An alignment aid for vhf receivers", J R Compton, G4COM. *Rad Com,* June 1976, p36.

[5] The traditional method of measuring nf with a thermionic noise diode is to measure the diode current corresponding to a 3dB rise in (s+n)/n ratio. The noise power from the generator is then taken to equal the receiver's own noise-floor power level. But this is only true in the absence of other noise sources. The resistive output circuit of the noise or signal generator has a noise temperature of its own, so its thermal noise will add to that of the receiver and will tend to mask any changes in the true (s+n)/n ratio. Thus the traditional "3dB" thermionic diode technique is not accurate for low-noise front-ends; it would need to be entirely reworked in terms of noise temperature.

[6] "An extremely low-noise 96MHz crystal oscillator for uhf/shf applications", B Neubig, DK1AG. *VHF Communications*, winter 1981.

[7] "A two-tone test generator for the two metre band", M Ploetz, DL7YC, in *VHF/UHF Technik* (Berliner DUBUS Gruppe, 1978).
UHF-Unterlage Teil 3, ed K Weiner, DJ9HO (in German) contains designs for both 144MHz and 432MHz.

[8] *Solid State Design for the Radio Amateur* (see [2]), p197. The bipolar feedback amplifiers shown following the diode doublers are not suitable for uhf. Other more suitable vhf/uhf high-level, low-noise, Class A amplifiers may be substituted.

[9] "Intermodulation measurement techniques in mixers and amplifiers", P E Chadwick. Application Note AN1009, Plessey Semiconductors Ltd. □